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(71) Applicant(s):
Hewlett-Packard Company
(Incorporated in USA - Delaware)
3000 Hanover Street, Palo Alto,
California 94304, United States of America

(72) Inventor(s):
Filippo Tosato
Paola Bisaglia

(74) Agent and/or Address for Service:
Richard Anthony Lawrence
Hewlett-Packard Limited, IP Section,
Filton Road, Stoke Gifford, BRISTOL,
BS34 8QZ, United Kingdom

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INT CL[7] **H03M** 13/41 13/45, **H04L** 1/00 25/02 27/26 27/34 27/38
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(54) Abstract Title: **Calculating an estimate of bit reliability in a OFDM receiver by multiplication of the channel state modulus**

(57) A scheme for simplifying the computational complexity of calculating log likelihood ratios (LLR) for soft output demapping is provided. The scheme can be implemented inside a receiver in an OFDM system using multi-level modulation whereby calculation of LLR ratios is accomplished using only multiplication of the channel state (CSI) modulus by d_K (which represents the half distance between partition boundaries in a QAM constellation).

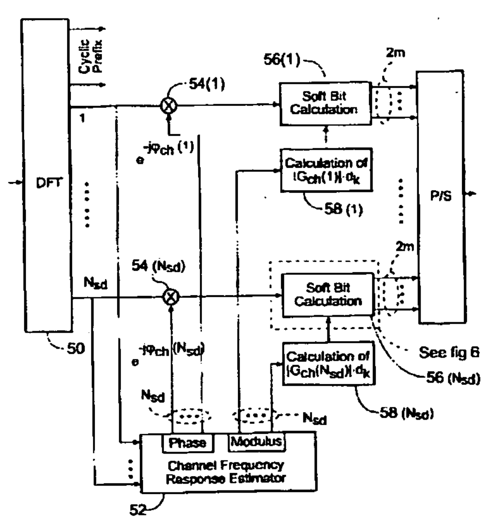

Fig. 5

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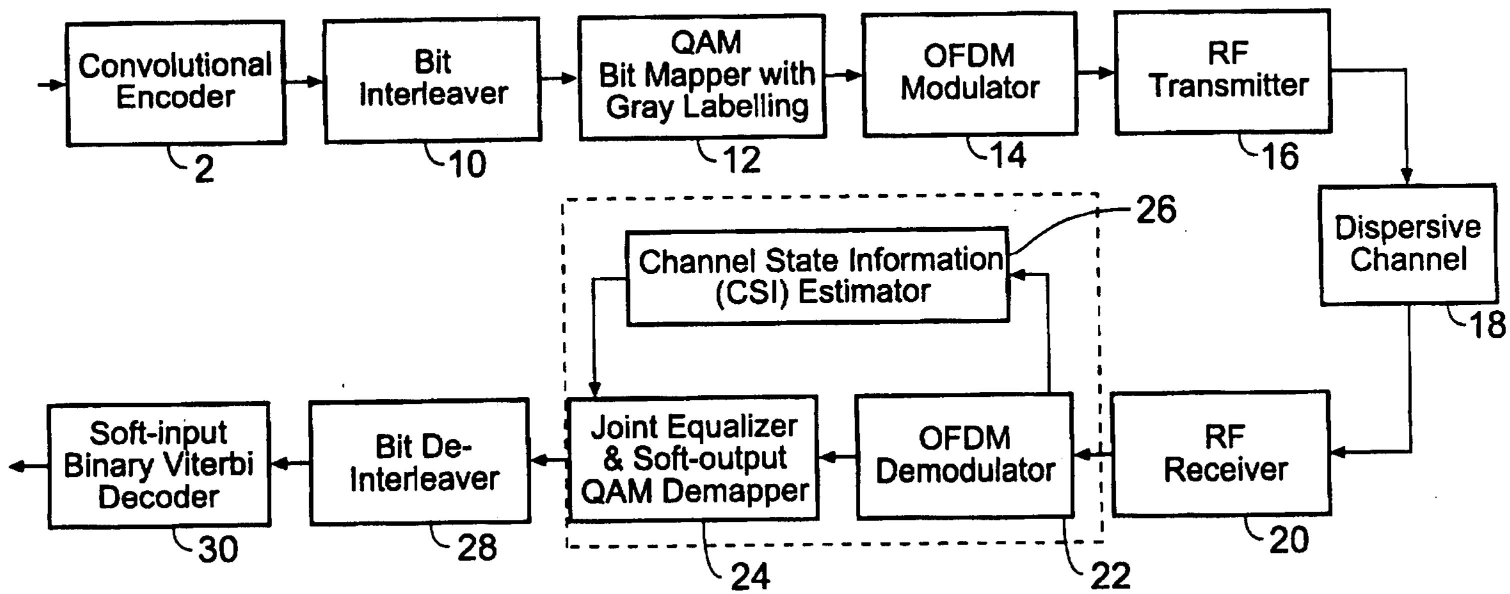

Convolutional Encoder
2
Bit Interleaver
10
QAM Bit Mapper with Gray Labelling
12
OFDM Modulator
14
RF Transmitter
16
Dispersive Channel
18
RF Receiver
20
OFDM Demodulator
22
Channel State Information (CSI) Estimator
26
Joint Equalizer & Soft-output QAM Demapper
24
Bit De-Interleaver
28
Soft-input Binary Viterbi Decoder
30

Fig. 1

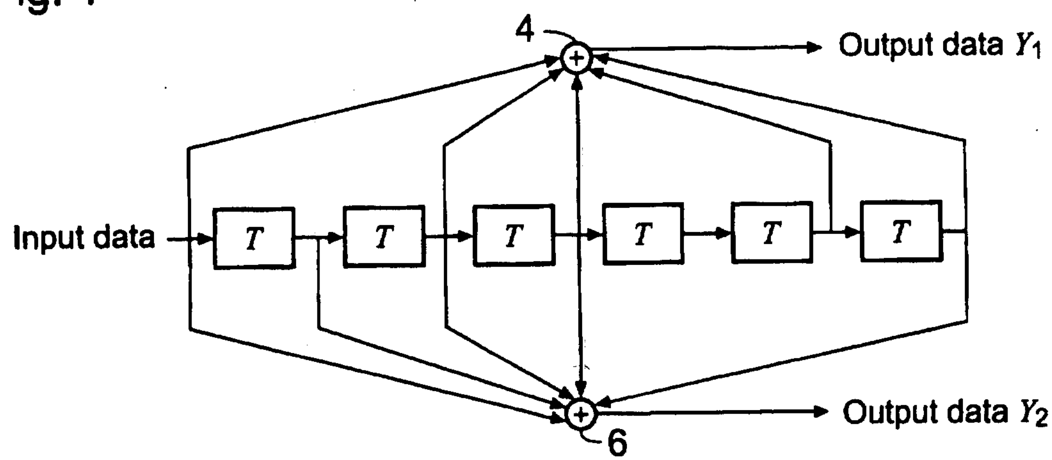

4
Output data Y_1
Input data
T
T
T
T
T
T
Output data Y_2
6

Fig. 2

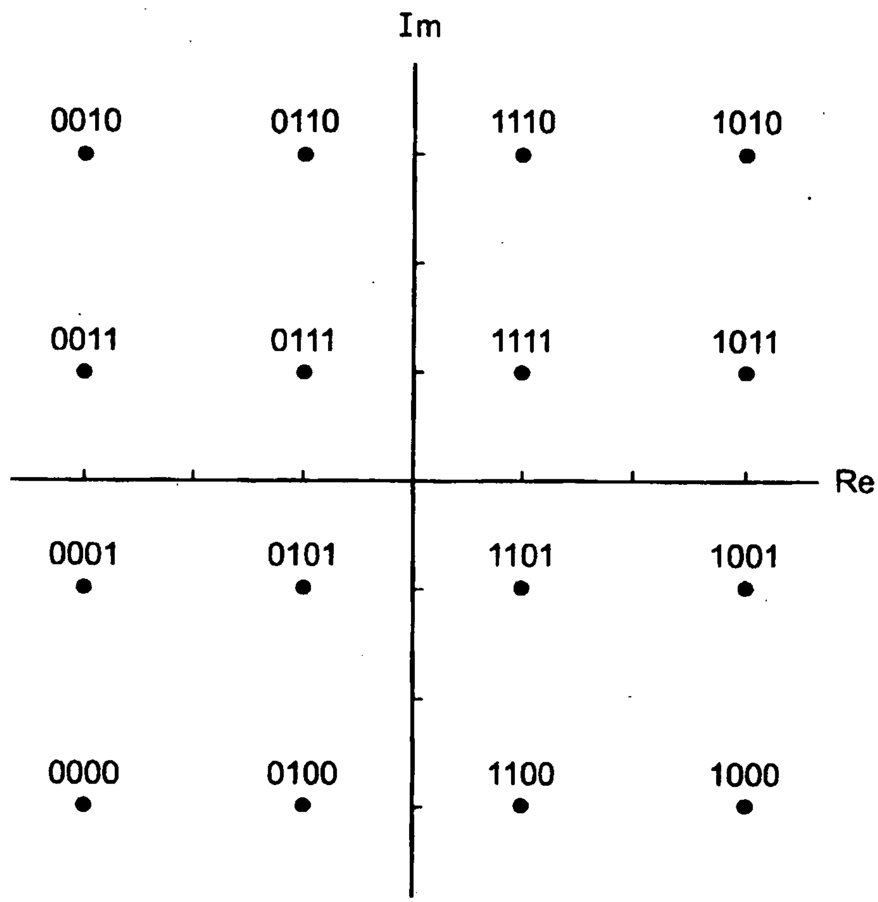

Im
0010
0110
1110
1010
0011
0111
1111
1011
Re
0001
0101
1101
1001
0000
0100
1100
1000

Fig. 3

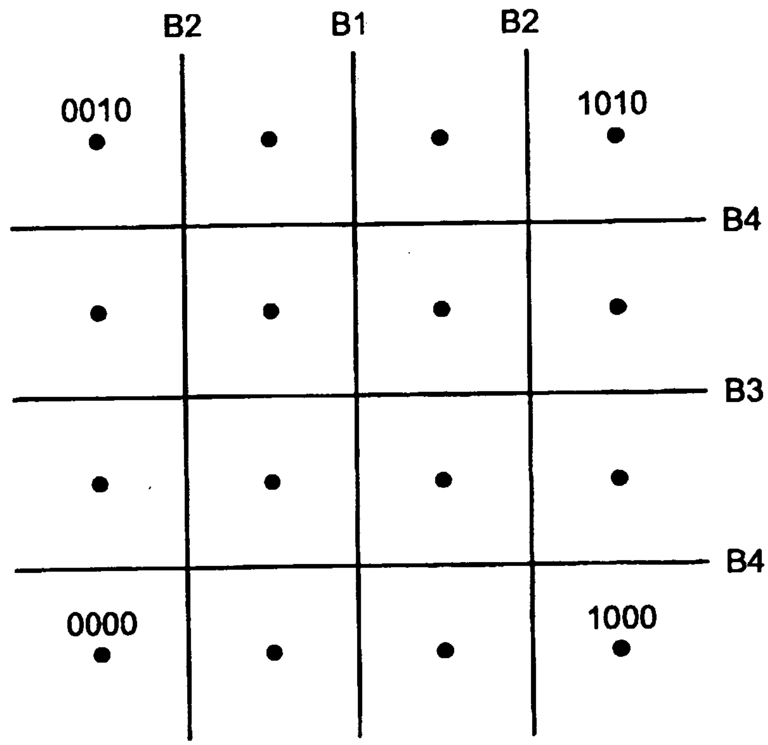


Fig. 4a

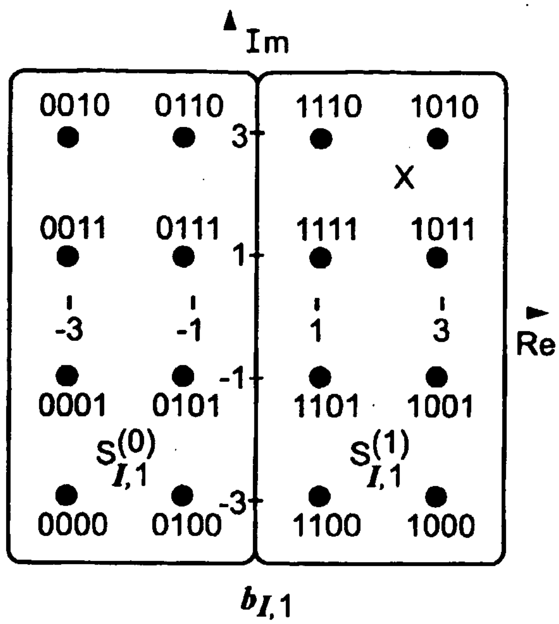


Fig. 4b

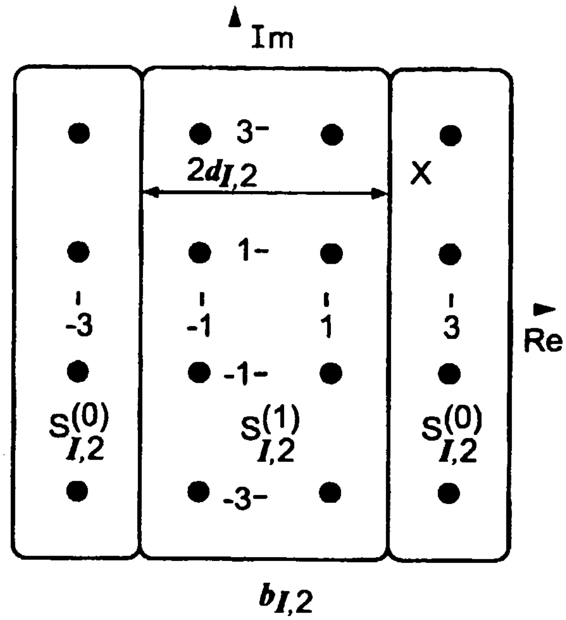


Fig. 4c

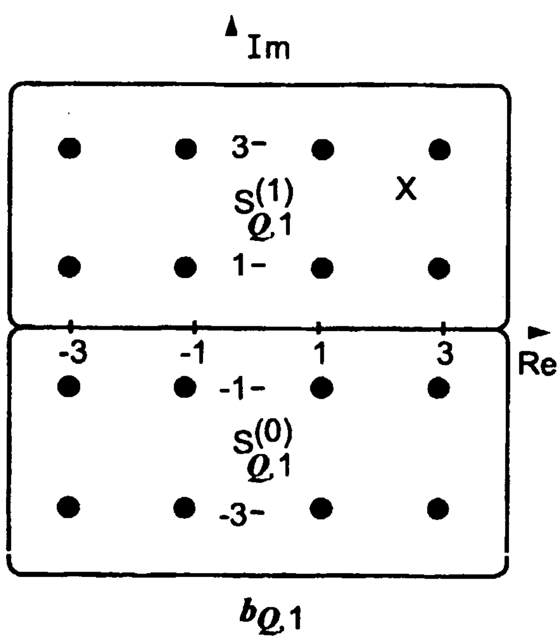


Fig. 4d

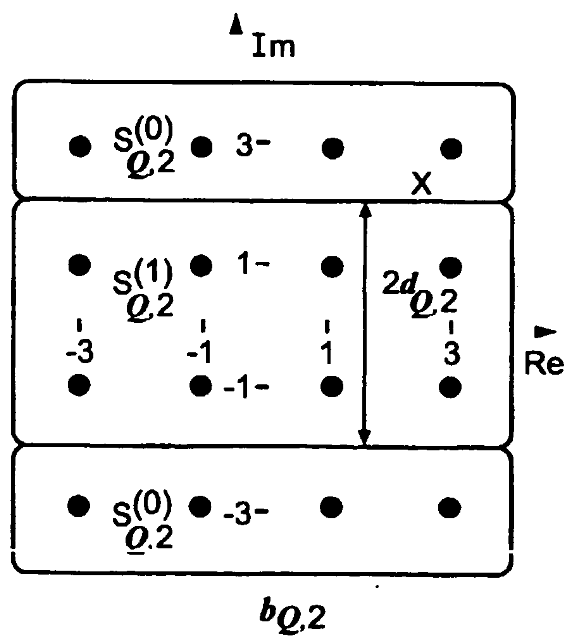
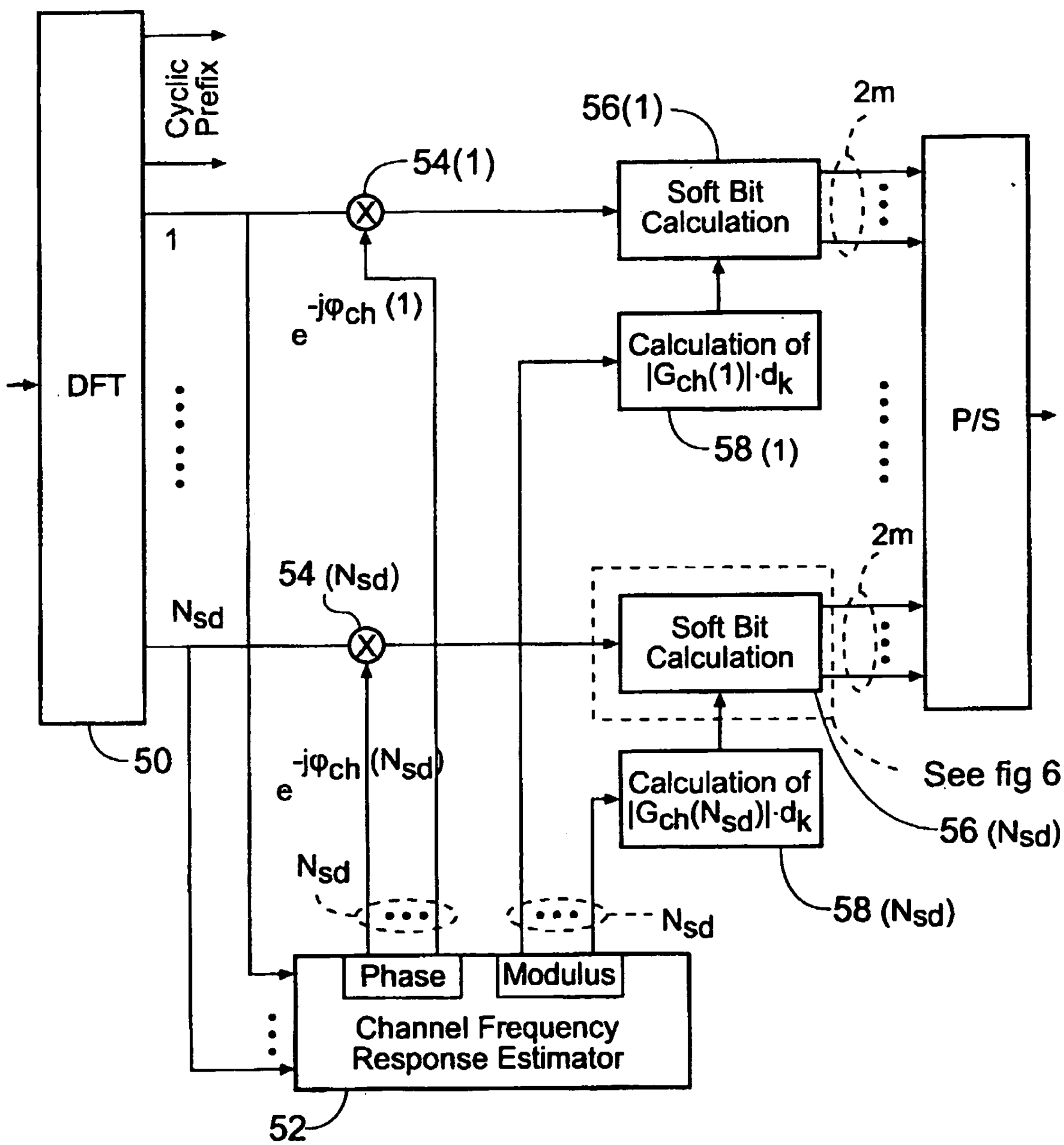


Fig. 4e



DFT
Cyclic Prefix
1
N_{sd}
50
54(1)
54 (N_{sd})
$e^{-j\varphi_{ch}(1)}$
$e^{-j\varphi_{ch}(N_{sd})}$
N_{sd}
Phase
Modulus
Channel Frequency Response Estimator
52
N_{sd}
56(1)
Soft Bit Calculation
Calculation of $|G_{ch}(1)| \cdot d_k$
58 (1)
Soft Bit Calculation
Calculation of $|G_{ch}(N_{sd})| \cdot d_k$
58 (N_{sd})
56 (N_{sd})
See fig 6
2m
2m
P/S

Fig. 5

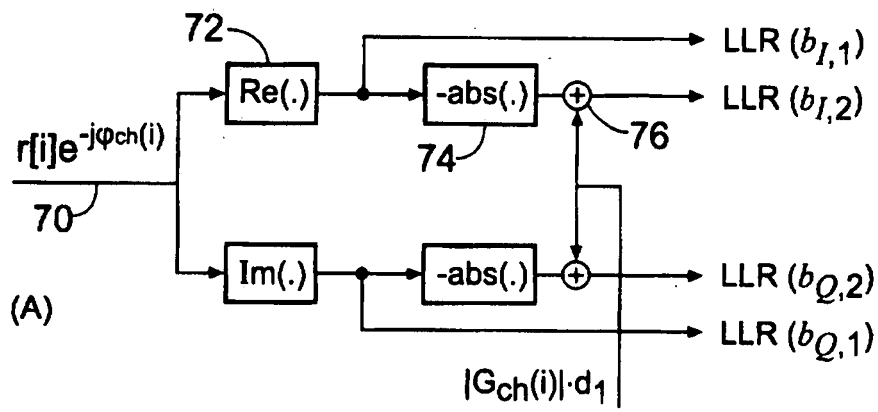

72
Re(.)
-abs(.)
LLR ($b_{I,1}$)
LLR ($b_{I,2}$)
$r[i]e^{-j\varphi_{ch}(i)}$
74
76
70
Im(.)
-abs(.)
LLR ($b_{Q,2}$)
LLR ($b_{Q,1}$)
(A)
$|G_{ch}(i)| \cdot d_1$

Fig. 6

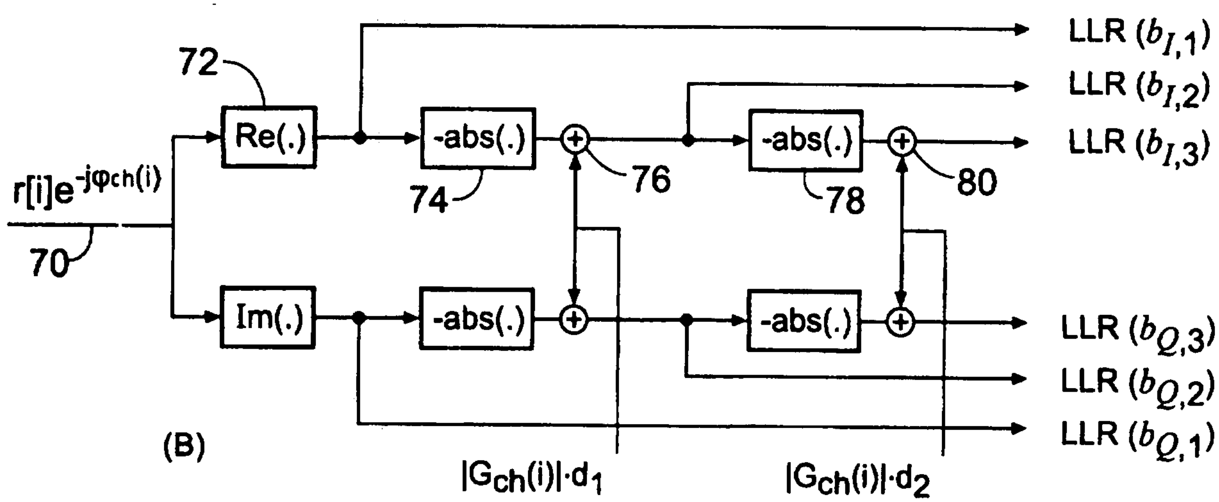

72
Re(.)
-abs(.)
-abs(.)
LLR ($b_{I,1}$)
LLR ($b_{I,2}$)
LLR ($b_{I,3}$)
$r[i]e^{-j\varphi_{ch}(i)}$
74
76
78
80
70
Im(.)
-abs(.)
-abs(.)
LLR ($b_{Q,3}$)
LLR ($b_{Q,2}$)
LLR ($b_{Q,1}$)
(B)
$|G_{ch}(i)| \cdot d_1$
$|G_{ch}(i)| \cdot d_2$

Fig. 7

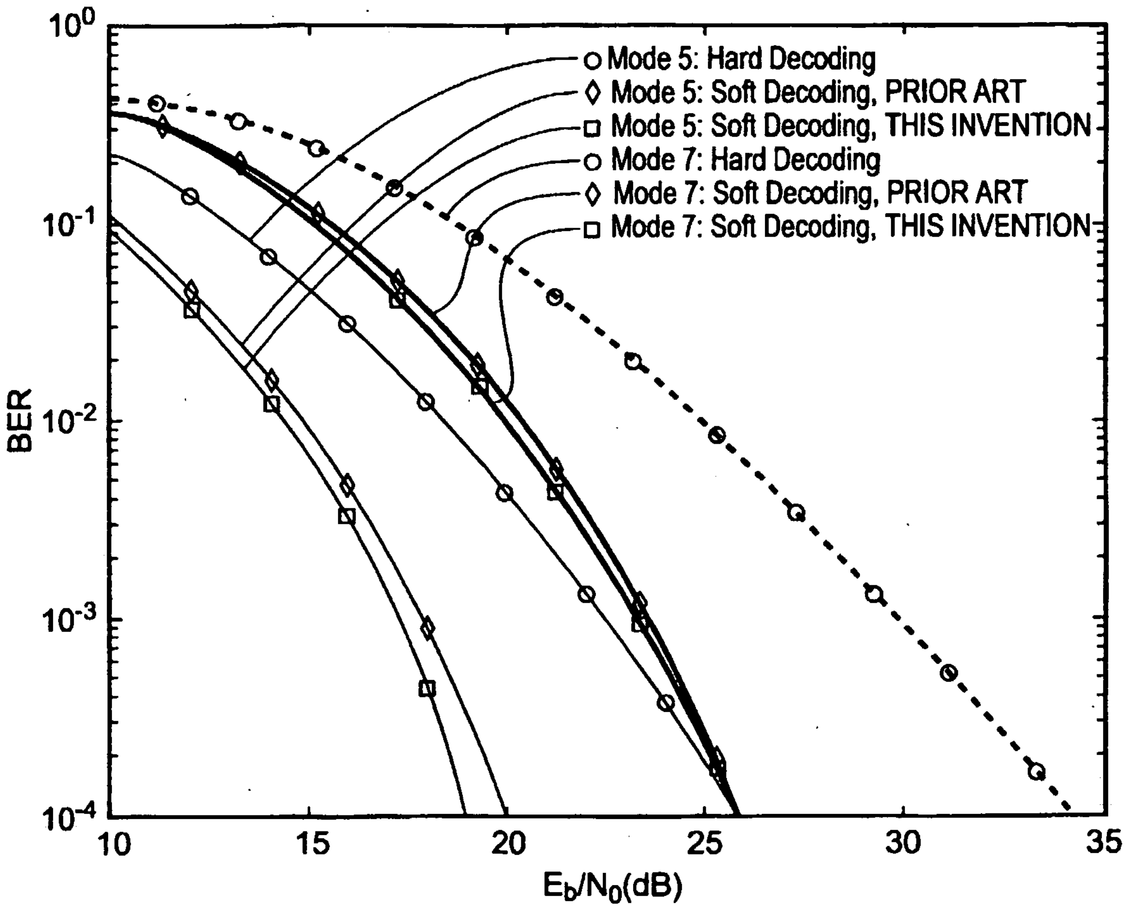

Mode 5: Hard Decoding
Mode 5: Soft Decoding, PRIOR ART
Mode 5: Soft Decoding, THIS INVENTION
Mode 7: Hard Decoding
Mode 7: Soft Decoding, PRIOR ART
Mode 7: Soft Decoding, THIS INVENTION
BER
10^0
10^{-1}
10^{-2}
10^{-3}
10^{-4}
10
15
20
25
30
35
E_b/N_0(dB)

Fig. 8

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A METHOD OF CALCULATING AN ESTIMATE OF RELIABILITY OF BIT INFORMATION IN A OFDM RECEIVER AND A RECEIVER OPERATING IN ACCORDANCE WITH THE METHOD

Background to the Invention

The present invention relates to a method of estimating a reliability measurement for a bit to be decoded in a multi-level transmission scheme based on bit-interleaved coded orthogonal frequency division multiplexing. The method significantly reduces the computational task of the receiver. The present invention also relates to a receiver and demapper operating in accordance with the method.

A significant amount of effort has been expended in the development of high data rate digital communication. Such communication technologies have brought about wireless local area networks, such as those defined in IEEE802.11a and HIPERLAN/2.

Reducing the computational complexity in the receivers of such a network would be beneficial for implementation of such systems because of the limited resources available in many low cost commercial devices.

Figure 1 schematically illustrates a communication system wherein the present invention can be applied. The transmitter part is in accordance with the HIPERLAN/2 and 802.11a standards. In both of these standards different QAM modulation formats are used, such as BPSK, QPSK, 16 QAM and 64 QAM.

In general a first user (who may be human or machine) wishes to transmit a data stream to a second user. The bits in the data stream are firstly convolutionally encoded by an encoder 2 to implement forward error correction. Schemes for convolutional encoding are well known to the person skilled in the art and hence only a brief description is required here. In general convolutional encoding is performed by shifting the data through a chain of delay elements usually implemented as a shift register. In the case of the HIPERLAN/2 and 802.11a standards the convolutional code has a rate ½ and a constraint length of seven. The chain is, as shown in Figure 2 tapped at various points along it and the tapped signals

are summed together at summers 4 and 6 which are implemented as modulo 2 adders, which are often physically realised as a cascade of exclusive - OR gates.

The arrangement shown in Figure 2 has six delay elements T and data can also be tapped off at the input to the first element. The encoder shown has, in effect, seven snapshots of the input string and is therefore described as having a constraint length of 7.

The encoder produces two outputs Y_1 and Y_2 in accordance with the series of connections and taps that are formed within it. Working from the input to the output, the first output is represented by the first, the third, fourth, sixth and seventh taps, or 1011011. The second output Y_2 is represented by 1111001. Thus the code generator for this encoder is [133, 171] in octal format. It should be noted that since the convolutional code has rate ½ the data entering the encoder is a scalar bit stream, whilst the data leaving the encoder is a stream of binary bit pairs. The data output of the encoder 2 is sent to a bit interleaver 10 so as to spread the bits so that potential errors caused by channel degradation are seen as independent at the receiver. The interleaved bits are then converted into quadrature - amplitude modulated (QAM) signals at a QAM mapper 12.

In other words sequences of 2m bits are assigned to one of the $M=2^{2m}$ points in the QAM constellation. The mapping is performed as a Gray coding such that sequences of bits associated with adjacent symbols only differ by one bit.

An exemplary 16-QAM constellation is shown in Figure 3.

The complex (real and imaginary parts) signals are then fed to an orthogonal frequency division multiplexing (OFDM) modulator 14. OFDM is a multi-carrier modulation scheme that subdivides the frequency spectrum that it uses into a plurality of channels. The scheme gives good performance over a dispersive transmission path. The data stream is split into N_{SD} lower rate data streams that are transmitted over N_{SD} sub-carriers. The symbol duration increases for the parallel sub-carriers and hence the transmission scheme is more robust in the presence of multi-path interference.

Finally the signal is transmitted by a radio frequency transmitter 16 such that the radio signal can propagate via a transmission channel 18 to a receiver 20.

The channel may be dispersive and noisy.

Consider the performance of a generic i_{th} sub-carrier used in the OFDM scheme. The i_{th} channel carries a series of QAM symbols. If we consider just one of these, then

$$a[i] = a_I[i] + ja_Q[i] \tag{1}$$

Where

$a[i]$ represents a symbol in the i_{th} sub-carrier

$a_I[i]$ represents the real component of the symbol

$a_Q[i]$ represents the imaginary component of the symbol.

j represents the imaginary operator.

Each symbol represents a convolutionally encoded bit sequence given by $\{b_{I,1};\ b_{I,2}.....;\ b_{I,m},\ b_{Q,1};\ b_{Q,2};\ ....\ b_{Q,m}\}$, as shown in Figure 3 for a 16 QAM constellation..

The output of the receiver 20 is provided to an OFDM demodulator 22 whose output is fed to an equalizer and soft-output QAM demapper 24. Assuming that the cyclic prefix, introduced by the OFDM modulator 14, completely eliminates ISI (Inter OFDM Symbol Interference) and ICI (Inter Channel Interference) and that the channel estimate is error free, the received equalized signal in the i_{th} sub-carrier is given by:

$$y[i] = r[i]/G_{ch}(i) = a[i] + w[i]/G_{ch}(i) \tag{2}$$

Where

$r[i]$ is the received signal before equalization at the output of the OFDM demodulator 22 in the i_{th} sub carrier,

$G_{ch}(i)$ is the Channel Frequency Response (CFR) coefficient (a complex number) in the i_{th} sub-carrier, and

$w[i]$ is the complex Additive White Gaussian Noise (AWGN).

The de-mapper 24 also receives information from a channel state information (CSI) estimator 26. The CSI estimator 26 attempts to deduce the effect of the transmission path on the channel.

In broad terms the channel both scales the QAM signals and rotates them in phase ie the QAM constellation gets rotated. The CSI estimator attempts to estimate the effect of the channel on the symbol. The CSI estimator can work on the assumption that channel is slowly time variant. However, the physical transmission scheme includes "preambles" which contain known sequences of symbols and these can be used to estimate the channel status. Thus individual phase and modulus estimates can be derived for each channel on the basis of identifying the preamble.

The output of the demapper 24 is provided to a de-interleaver 28 and to a convolutional decoder 30. The Viterbi algorithm is a widely used method of decoding convolutional codes. The algorithm searches the possible code words of the convolutional code and detects the one that is most likely to have generated the received sequence. The search procedure steps through the code trellis and for each path along the trellis computes a metric which quantifies the discrepancy between the received sequence and the possible coded sequence. If the information associated to bits fed into the decoder is *hard* (ie binary, a sequence of -1's and 1's) the decoder is called *hard decision* decoder. Alternatively, if the information is *soft*, consisting of a hard decision (the sign) and a confidence level, or reliability (the magnitude), that represents how much confidence there should be in the hard-decision, the decoder is called *soft-decision* decoder. It's well known that, soft decision Viterbi decoding can give significant gain over hard decoding at the expense, however, of a greater computational complexity. In order to implement a soft-decision Viterbi decoder the demapper, which precedes the Viterbi decoder, needs to deliver soft information associated to the bits.

Soft decision demapping for bit interleaved coded modulation (BICM) signals with BPSK or QPSK modulations is straightforward as the soft bit information, before being weighted by the Channel State Information (CSI) is simply given by the received signals for BPSK and by their in-phase and quadrature components for QPSK. Therefore, in the following

discussion consideration will be given to the higher modulation formats, for which soft detection requires much more computational effort.

In the literature, two different approaches can be found to calculate the soft information for BICM signals, with multi-level modulations.

The first prior art mechanism for bit interleaved coded modulation (BICM) schemes was disclosed by E. Zehavi, "8-PSK Trellis Codes for a Rayleigh Channel" IEEE Trans on Comm, Vol 40, pp 873 - 884, May 1992. The process starts by calculating sub-optimal bit metrics that are then used inside a Viterbi decoder for path metric computation.

For each bit $b_{I,K}$ and each bit $b_{Q,K}$ (where I and Q represent in-phase and quadrature parts, respectively, and K represents an index of the bit associated with the symbol where K is an integer in the range $1 \leq K \leq m$) the QAM constellation is split into two partitions of complex symbols.

These partitions are

$S_{I,K}^{(0)}$ having symbols with 0 in position I, K.

$S_{I,K}^{(1)}$ having symbols with 1 in the position I, K.

$S_{Q,K}^{(0)}$ having symbols with 0 in the position Q, K.

$S_{Q,K}^{(1)}$ having symbols with 1 in the position Q, K.

The bit metrics are given by

$$M_C(b_{I,K}) = |G_{CH}(i)|^2 \cdot \min_{a \in S_{I,K}^{(c)}} |y[i]-a|^2 \quad , \qquad c=0,1 \tag{3}$$

Finally the metrics are de-interleaved by a de-interleaver 28 and provided as an input to a Viterbi decoder 30.

The Viterbi decoder works according to a well known algorithm which need not be described in detail here. However various web sites give tutorials in Viterbi decoding, such as http://pweb.netcom.com/~chip.f/viterbi/algrthms2.html.

The convolutional encoder 2 functions as a state machine and the Viterbi decoder is furnished with a state map of the state machine showing which state to state transitions are allowed and which ones are disallowed.

In the second prior art approach the QAM symbols are first demodulated by a soft output de-mapper and passed to a soft-input Viterbi decoder, see M. Speth et al, "Low Complexity Space-Frequency MLSE for Multi-User COFDM", IEEE GLOBECOM '99, pp 2395 - 99, Dec. 1999.

In this approach the process seeks to de-map the received signal into soft bits which have the same sign as that provided by a hard decoder and whose magnitude indicates the reliability of the decision.

The soft bit information assigned to bit $b_{I,K}$ can be shown to be given by the log-likelihood ratio (LLR) of the hard decision on $b_{I,K}$ (see R. Pyndiah et al "Near Optimum Decoding of Product Codes", IEEE GLOBECOM 94, pp 339 - 43, Nov. - Dec., 1994) and can be approximated by

$$LLR(b_{I,K}) = \left|\frac{G_{CH}(i)}{4}\right|^2 \left\{ \min_{a \in S_{I,K}^{(0)}} |y[i]-a|^2 - \min_{a \in S_{I,K}^{(1)}} |y[i]-a|^2 \right\} \qquad (4)$$

$$LLR(b_{I,K}) \triangleq |G_{CH}(i)|^2 . D_{I,K} \qquad (5)$$

We define $S'^{(c)}_{I,k} \triangleq \Re\{S^{(c)}_{I,k}\}$ as the subset containing the real parts of the complex symbols of subset $S^{(c)}_{I,k}$, for $c = 0,1$. It can be shown that equation (4) can be rewritten in a simpler form,

$$LLR(b_{I,k}) = \frac{|G_{ch}(i)|^2}{4} \left\{ \min_{a_I \in S'^{(0)}_{I,k}} (y_I[i]-a_I)^2 - \min_{a_I \in S'^{(1)}_{I,k}} (y_I[i]-a_I)^2 \right\} \qquad (6)$$

Where the two minima are now taken over real values instead of complex symbols.

If this is, for convenience, explicitly evaluated for the 16 QAM symbols we have

$$D_{I,1} = \begin{cases} y_I[i] & |y_1[i]| \leq 2 \\ 2(y_I[i] - 1 & y_I[i] > 2 \\ 2(y_I[i] + 1) & y_I[i] < 2 \end{cases}$$

$$D_{I,2} = -|y_I[i]| + 2 \tag{7}$$

equivalent expressions hold for the quadrature components with "I" replaced by "Q".

It has been demonstrated in F. Tosato and P. Bisaglia "Simplified Soft-Output Demapper for Binary Interleaved COFDM with Application to HIPERLAN/2", IEEE ICC 2002, April-May, 2002, that using the approximate bit metrics $M_C(b_{I,K})$ in equation 3 for path metric calculation inside the Viterbi algorithm is equivalent to demodulating the signals into soft bit values according to equation 6 and then employing a soft Viterbi algorithm for decoding.

The formula for calculating the log likelihood ratio in equation 6 can be further approximated by calculating $|D_{I,K}|$(or indeed $|D_{Q,K}|$) as the distance of the received equalised signal $y[i]$ from the nearest partition boundary within the partitioned QAM space and assigning $D_{I,K}$ (or $D_{Q,K}$ as appropriate) the sign "+" or "-" according to which partition $y[i]$ falls in. The magnitude (or absolute value) is a measure of distance of the received symbol from the decision boundary.

Figure 4a illustrates the decision boundaries B1 to B4 for 16 QAM modulation and Figures 4b to 4e illustrate the resulting partitions in 16 QAM space. The 16 QAM constellation is used by way of an example. However the present invention can be applied to higher order constellations in a similar way.

Furthermore if we let $d_{I,K}$ and $d_{Q,K}$ denote half the distance between the partition boundaries B2 and B4 relative to bit $b_{I,K}$ and $b_{Q,K}$, respectively, then for the 16 and 64 QAM constellations with Gray mapping used in IEEE802.11a and HIPERLAN/2 it can be shown that

$$D_{I,K} \simeq \begin{cases} y_I[i], & K = 1 \\ -|D_{I,K-1}| + d_{I,K}, & K > 1 \end{cases} \tag{8}$$

In terms of computational complexity the prior art system of decoding the symbols, even with all of the simplifications and approximations invoked, is computationally complex.

To illustrate this consider the case of a burst transmission. In such a burst transmission the channel can be assumed to be time-invariant for the duration of the burst. Thus channel state estimation need only be performed once by the receiver at the beginning of each burst.

If we denote N_b the number of bits coded in the data burst and N_{SD} the number of sub-carriers ($N_b >> N_{SD}$) then LLR calculation using the formulae:

$$LLR(b_{I,K}) = |G_{CH}(i)|^2 D_{I,K}$$

$$LLR(b_{Q,K}) = |G_{CH}(i)|^2 D_{Q,K} \tag{9}$$

requires one real multiplication per coded bit, plus computation of N_{SD} squared modulus of complex values per physical data burst, which is equivalent to $2N_{SD}$ real multiplications.

Thus, the approximate LLR calculation requires $(N_b + 2N_{SD})$ real multiplications.

According to a first aspect of the present invention there is provided a method of estimating a measure of trust of data conveyed by a QAM symbol, wherein the measure of trust is calculated as a linear function of the modulus of a channel state estimation.

It is thus possible to reduce the number of calculations, and in particular multiplications, performed in the soft output de-mapping by using an estimate of likelihood derived as a function of $|G_{CH}(i)|$ rather than $|G_{CH}(i)|^2$.

Preferably the estimate of trust for a bit $b_{I,K}$ for K=1, where K represents a bit index within a complex symbol, is calculated as $\Re\{r[i]e^{-j\varphi_{ch}(i)}\}$, where $\Re$ represents the "real" part of a complex number, $r[i]$ is the received symbol in the i_{th} sub-carrier, and $e^{-j\varphi_{ch}(i)}$ represents the reciprocal of the phase response of an i_{th} transmission channel over which the symbol was transmitted.

Preferably the estimate of trust for a bit $b_{Q,K}$ is calculated as $\Im\{r[i]e^{-j\varphi_{ch}(i)}\}$ where $\Im$ represents the imaginary part of a complex number, for K=1.

Preferably the estimate trust of a bit $b_{I,K}$ is further calculated as $-|LLR(b_{I,K-1})| + (|G_{ch}(i)| \cdot d_{I,K})$ for $K > 1$ where $G_{ch}(i)$ represents the channel frequency response complex coefficient on an i_{th} channel and $d_{I,K}$ denotes a half distance between partition boundaries in QAM space, for $K > 1$, and the estimate of trust of a bit $b_{Q,K}$ is calculated as $-|LLR(b_{Q,K-1})| + (|G_{ch}[i]| \cdot d_{Q,K})$ where $d_{Q,K}$ denotes the half distance between partition boundaries in QAM space, for $K > 1$.

Preferably the estimate of trust is approximate log-likelihood ratio.

According to a second aspect of the present invention, there is provided an apparatus for performing the method according to the first aspect of the present invention.

The present invention will now further be described, by way of example only, with reference to the accompanying figures:

Figure 1 schematically illustrates a transmit path and a receive path for a bit interleaved coded modulation scheme;

Figure 2 schematically illustrates a convolutional encoder;

Figure 3 illustrates a square 16 QAM constellation;

Figure 4a illustrates the partition boundaries that are used to partition the 16 QAM space;

Figures 4b to 4e illustrate the resulting partitions for a 16 QAM space;

Figure 5 is a schematic diagram of a channel state estimator and soft output QAM demapper constituting an embodiment of the present invention;

Figure 6 schematically illustrates a soft decision block constituting an embodiment of the present invention for 16 QAM constellation;

Figure 7 schematically illustrates a soft decision block constituting an embodiment of the present invention for 64 QAM; and

Figure 8 is a comparison of decoding schemes, with the curves showing simulations for performance of a HIPERLAN/2 system.

It should be noted that, for simplicity, the present invention is described with respect to 16 QAM and 64 QAM, but it can be applied to larger constellations and similar demapping schemes can be derived.

Unlike single carrier systems in which all symbols are affected by the same signal to noise ratio (on average), a multi-carrier OFDM system of the type shown in Figure 1 is such that each individual carrier suffers from an individual signal to noise ratio. However it is clear to the person skilled in the art that data conveyed on channels having a high signal to noise ratio is *a priori* more reliable than data transmitted on channels suffering from a low signal to noise ratio. This additional information has, in the prior art, been encoded by weighting the LLR functions by the square modulus of the channel frequency response, which represents the channel state information.

However the inventors have realised that using the modulus of the channel frequency response coefficients instead of the square of the modulus for calculating "soft bit" information for use by a decoder results in only a slight performance loss in terms of bit error rate at a given signal strengths as represented by E_b/N_o, where E_b is the energy per information bit and N_o is the power spectral density of the noise.

This approximation does, however, allow for a computationally efficient implementation of a one tap equaliser and LLR calculation subsystem. Thus, the inventors have realised that sub-optimum soft input Viterbi decoding of a binary interleaved coded OFDM signal can be achieved with little additional complexity compared to realising the same operation using hard decoding instead.

By applying this approximation to the prior art scheme for calculating the LLRs, the following equations are obtained.

$$LLR(b_{I,K}) = \begin{matrix} \Re\{r[i]e^{-j\varphi_{ch}(i)}\}, & K=1 \\ -|LLR(b_{I,K-1})| + |G_{ch}(i)| \cdot d_{I,K} & K \geq 1 \end{matrix}$$

$$LLR(b_{Q,K}) = \begin{cases} \Im\{r[i]e^{-j\varphi_{ch}(i)}\}, & K=1 \\ -\left|LLR(b_{Q,K-1})\right| + |G_{ch}(i)| \cdot d_{Q,K} & K \geq 1 \end{cases} \tag{10}$$

Where $\qquad G_{ch}(i) = |G_{ch}(i)|e^{j\varphi_{ch}(i)}$

Similar results hold for other Gray labelling patterns to that shown in Figure 3.

Thus, compared with the prior art calculations of LLR, it can be seen that inside the LLR expression only the thresholds $d_{I,K}$ and $d_{Q,K}$ are scaled by the coefficients that convey the channel state information.

As a consequence, the following scheme can be adopted for joint OFDM signal equalisation and LLR computation. A block diagram for implementing the scheme is illustrated in Figure 5.

The incoming OFDM signal from a receiver is converted into individual data channels 1 to N_{SD} by a Fourier transform block 50. The phase of the signal is then equalised/corrected. This is done by sending each one of the channels to an input of a channel frequency response estimator 52 which estimates the phase of each one of the channels and thereby produces a phase correction signal $e^{-j\varphi_{ch}(i)}$ for each channel i, where i is an integer in the range $1 \leq i \leq N_{SD}$. The channel state estimate is done once per physical burst, at the beginning thereof.

The phase corrections are applied to each of the channels via respective multipliers 54(1) to 54(N_{SD}). The phase equalised channel signals are then supplied to first inputs of respective soft bit calculators 56(1) to 56(N_{SD}).

The channel frequency response estimator 52 also estimates the modulus of the signal strength in each one of the channels, and this information is supplied to a threshold calculator 58(1) to 58(N_{SD}) associated with each individual channel which calculates the threshold values $|G_{ch}(i)| \cdot d_{I,K}$ and $|G_{ch}(i)| \cdot d_{Q,K}$ where i represents the channel number. These values are then passed to the respective soft bit calculators 56.

The specific implementation of the soft bit calculators is shown in Figure 6, for a 16 QAM constellation and in Figure 7 for a 64 QAM constellation.

For the arrangement shown in Figure 6, the received phase equalised signal is received at an input 70 and supplied to a first analyser 72 which calculates the real component of the signal. The output of the first analyser represents bit value $LLR(b_{I,1})$. The value $LLR(b_{I,1})$is further passed through an absolute value former 74 which calculates and negates the absolute value of $LLR(b_{I,1})$. The output of the absolute value former 74 is then added to $|G_{ch}(i)| \cdot d_I$ by an adder 76 to yield an output $LLR(b_{I,2})$. A similar process is implemented on the imaginary component of the input signal to yield $LLR(b_{Q,1})$and $LLR(b_{Q,2})$.

The arrangement shown in Figure 7 is similar to that shown in Figure 6 and like parts are denoted by like reference numerals. The additional features are that the output $LLR(b_{I,2})$ is made available to a further adder absolute value former and negator 78 whose output is then added to $|G_{ch}(i)| \cdot d_2$ by a further 80 to yield $LLR(b_{I,3})$. Similar components are provided in the "imaginary" signal path for calculating $LLR(b_{I,3})$.

Comparing the complexity of this scheme in a burst mode with the prior art, it is now necessary to calculate N_{SD} real multiplications per transmission burst in the case of 16 QAM and $2N_{SD}$ for 64 QAM because only the thresholds d_K need to be scaled by the cannel state information coefficients.

However, these multiplications are not actual multiplications because for square QAM constellations d_K are powers of two, so in fact only bit-wise shifts are needed for threshold scaling.

It is, however, necessary to calculate N_{SD} moduli of the complex values per physical burst to calculate the channel state information values. This requires $2N_{SD}$ real multiplications and N_{SD} square roots to be formed.

Comparing the computational workload:

Present invention = $2N_{SD}$ real multiplications $+N_{SD}$ square roots.

Prior art = $2N_{SD}$ real multiplications $+N_b$ real multiplications.

The computational advantage comes because $N_b >> N_{SD}$ and in practice N_{SD} is 48 in HIPERLAN/2 whereas N_b is likely to be several thousand.

The performance of the present invention compared to the computationally complex prior art is shown in Figure 8. For a given bit error rate the degration in sensitivity is only a fraction of a dB for 16 QAM and is negligible for 64 QAM at BER 10^{-4}. The approximation gets tighter for larger constellation size, where more calculations are saved by the approximate demapping. The graph compares hard decoding, prior art soft decoding and soft decoding according to the present invention in modes 5 and 7 of HIPERLAN/2 using 16 and 64 QAM respectively.

CLAIMS

1. A method of estimating a measure of trust of data conveyed by a QAM symbol, wherein the measure of trust is calculated as a linear function of the modulus a channel state estimation.

2. A method as claimed in claim 1, wherein the measure of trust is a log-likelihood ratio.

3. A method as claimed in claim 2 where the $LLR(b_{I,K})$, for K=1, is calculated as $\Re\{r[i]e^{-j\varphi_{ch}(i)}\}$, where $\Re$ represents the "real" part of a complex number, $r[i]$ is the received symbol in the i_{ch} channel, and $e^{-j\varphi_{ch}(i)}$ represents the reciprocal of the phase response of the i_{th} transmission channel over which the symbol was transmitted, where K represents a bit index within a complex symbol.

4. A method as claimed in claim 3, where the $LLR(b_{Q,K})$ for K=1 is calculated as $\Im\{r[i]e^{-j\varphi_{ch}(i)}\}$ where $\Im$ represents the imaginary part of a complex number.

5. A method as claimed in claim 3, where the $LLR(b_{I,K})$ is calculated as $-|LLR(b_{I,K-1})|+(|G_{ch}(i)|\cdot d_{I,K})$ for $K>1$ where $G_{ch}(i)$ represents the channel frequency response complex coefficient on an i_{th} channel and $d_{I,K}$ denotes a half distance between partition boundaries in QAM constellation.

6. A method as claimed in claim 3, where the $LLR(b_{Q,K})$ is calculated as $-|LLR(b_{Q,K-1})|+(|G_{ch}[i]|\cdot d_{Q,K})$ for $K>1$ where $d_{Q,K}$ denotes the half distance between partition boundaries in QAM constellation.

7. A method as claimed in claim 3, wherein thresholds $d_{I,K}$ are scaled by the amplitude of the channel state estimator coefficients $G_{ch}(i)$.

8. A method as claimed in claim 7, in which computational complexity is reduced by using the modulus of the estimated channel state estimator coefficients $|G_{ch}(i)|$ instead of $|G_{ch}(i)|^2$ as channel state information inside a soft output demapper.

9. A method as claimed in claim 1, wherein the QAM symbol is a symbol in a OFDM transmission scheme using multi-level modulation.

10. A soft output demapper operating in accordance with the method claimed in claim 1.

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UK Patent Office collections, including GB, EP, WO & US patent specifications, in:

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Other: Online: EPODOC, JAPIO, WPI, Inspec, Full Text Databases

Documents considered to be relevant:

Category	Identity of document and relevant passage		Relevant to claims
X	US 2002037057 A1	[KROEGER] See para 14	1 at least
X	US 6078626 A	[RAMESH] See equations in columns 5 and 6	1 at least
X	US 5134635 A	[HONG] See fig 2a and columns 3 and 4	1 at least

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